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THE FREQUENCY DIFFERENCE TRANSLATOR

AN IMPROVEMENT IN AIRBORNE TECHNIQUES FOR MEASURING THE EARTH'S TOTAL MAGNETIC FIELD GRADIENT

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SUMMARY

Improvements in Airborne techniques for the measurement of the earth's total magnetic field gradient using optically pumped magnetometers are discussed. These techniques involve fast and accurate measurements of small frequency differences and are applicable to a large class of transducers which have for an output a frequency varying with the input variable. Description and design details are given of a newly developed analog instrument which allows a continuous read-out of frequency difference with a resolution of better than 0.01 Hz.

RÉSUMÉ

L'auteur discute des améliorations des techniques aériennes pour la mesure du gradient du champ magnétique total de la terre en utilisant des magnétomètres à pompage optique. Ces techniques comportent des mesures rapides et précises de petites différences de fréquence et sont applicables à une vaste classe de transducteurs qui ont comme sortie une fréquence variant en fonction de la variable d'entrée. L'auteur fournit la description et les détails de conception d'un instrument analogique de production récente qui permet une lecture continue de la différence de fréquence avec une précision supérieure à 0.01 Hz.

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SYMBOLS

Symbol	Definition
A	frequency modulation index
B	constant
DLD	delay line frequency discriminator
$\bar{D}_n$	difference of two digital frequency readings (Hz)
F, F_j	modulating frequency in frequency modulation (Hz)
FDT	frequency difference translator
Hz	Hertz
K	frequency multiplication constant
MHz	10^6 Hertz
N	accumulated count in digital counter
PLL	phase locked loop
S(F)	digital counter response to frequency modulated input
T	time interval, counter gate time (seconds)
V_R	reference voltage (volts)
f, f_j	frequency (Hz)
f(t), $f_j(t)$	time varying frequency (Hz)
f_L	Larmor frequency (Hz)
f_R	reference frequency (Hz)
$\bar{f}$	average value of frequency (Hz)
$\bar{f}_n$	average value of frequency in n^{th} sample (Hz)
mV	10^{-3} Volts
kHz	10^3 Hertz
o,k,m,p,q	constants
n	integer
t	time (seconds)

SYMBOLS (Cont'd)

Symbol	Definition
t_{on}	time when flip-flop Q output is high (seconds)
t_{off}	time when flip-flop Q output is low (seconds)
v_o	output voltage (Volts)
ϕ_M	measured total magnetic field (γ)
$\phi_M(t)$	time varying measured total magnetic field (γ)
$\phi_E(t)$	earth's total magnetic field (γ)
ϕ_j	amplitude of interfering magnetic field
$\phi_j(t)$	interfering magnetic field
Δf	frequency difference (Hz)
$\Delta \bar{f}$	difference of average value of frequencies (Hz)
α	frequency proportionality coefficient (Hz/γ)
γ	unit of magnetic field $1\gamma = 10^{-9}$ Tesla
τ	time delay (seconds)
ψ	phase angle (radians)
$\hat{\psi}(t)$	random component of phase angle (radians)
θ	fraction of frequency difference Δf, $0 \leqslant \theta \leqslant 1$
ω_1, ω_2	angular velocity (radians/second)

THE FREQUENCY DIFFERENCE TRANSLATOR
AN IMPROVEMENT IN AIRBORNE TECHNIQUES FOR MEASURING THE EARTH'S MAGNETIC FIELD GRADIENT

1.0 INTRODUCTION

An accurate airborne measurement of the earth's total magnetic field and its gradient poses numerous technical difficulties. To measure the gradient component along a line, instantaneous measurements are required at two proximate locations; then the difference of the measurements has to be computed and divided by the distance between the two locations. The gradient is more sensitive to close magnetic sources, while distant sources affecting both measuring instruments similarly will have a much smaller effect. The gradient, therefore, can be indicative of an ore body near the earth's surface or even the presence of a submarine.

Recently, there has been a surge of interest in airborne methods of measuring the gradient of the earth's total magnetic field. A considerable amount of experimental data has been obtained from the magnetometer-equipped Convair 580 and its predecessor the North Star aircraft, both operated by the Flight Research Laboratory of the National Aeronautical Establishment. Each of these aircraft used optically pumped cesium-vapour magnetometers; their output is a frequency proportional to the total magnetic field. The measurement of the gradient components is, therefore, reduced to the measurement of a frequency difference. To facilitate the measuring and processing of total magnetic field gradient, an analog instrument with high resolution has been specially developed by the Flight Research Laboratory.

This paper discusses a few presently used conventional measurement methods and their limitations when used with optically pumped magnetometers. It also discusses the design principles, the advantages of the newly-developed analog instrument and its current and proposed applications.

2.0 MAGNETIC FIELD MEASURING TECHNIQUES

The following discussion of measuring techniques is applicable to a large class of transducers which have for an output a frequency varying with an input variable. To this class of transducers belong optically pumped and nuclear resonance (NMR) magnetometers (Refs. 1, 2, 3, 4). The emphasis here is on optically pumped cesium vapour magnetometers.

It will be recalled that the earth's magnetic field is a vector field. The optically pumped cesium-vapour magnetometer measures only the magnitude of the earth's magnetic vector irrespective of its orientation (some magnetometers incorporate an angular positioning system aligning the magnetometer parallel to the earth's magnetic vector). Thus the magnitude of the earth's magnetic vector forms a scalar field referred to as the earth's total magnetic field. The future discussion relates to total magnetic fields which may originate from various magnetic dipoles which are summed with that of the earth.

In order to discuss methods of measuring the magnetic field difference it is necessary to describe how the magnetic field is measured and the limitations of the measurement process.

The output of a magnetometer is a frequency, proportional to the magnetic field ϕ_M, called the Larmor frequency f_L with

$$f_L = \alpha\phi_M \qquad (1)$$

where $\alpha \cong 3.5$ Hz/γ* for cesium and ϕ_M is in gammas (γ) (Ref. 5).

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* $1\gamma = 10^{-9}$ Tesla

The time varying magnetic field $\phi_M(t)$ as seen by the airborne magnetometer can be considered as the sum of the earth's field $\phi_E(t)$ and various interfering fields $\phi_j(t)$

$$\phi_M(t) = \phi_E(t) + \Sigma\phi_j(t) \tag{2}$$

To measure the magnetic field the Larmor frequency has to be measured. Usually the Larmor frequency is counted over a certain fixed time interval with a digital frequency counter. The accumulated count will correspond to the average value of the field over this time interval; the process is that of integration with respect to time. The averaging time must be long enough to reduce the effect of high frequency interference; but, at the same time, it must be short enough to follow the field changes with time such as those caused by the movement of the aircraft through the variable earth's field.

The digital counter used to measure the frequency is one of the main building blocks of the instrumentation and its behaviour will be briefly described. The frequency response of the counter can be calculated as follows: assume an instantaneous frequency f(t) of a frequency-modulated signal,

$$f(t) = f_o + A\sin 2\pi Ft \tag{3}$$

where $A < f_o$ and F is the frequency of modulation.

The average value $\bar{f}_n$ of f(t) is corresponding to the nth sample of duration T and it is defined as:

$$\bar{f}_n = \frac{1}{T}\int_{nT}^{(n+1)T}(f_o + A\sin 2\pi Ft)dt = f_o + A\sin[2\pi F(n+\tfrac{1}{2})T]\frac{\sin\pi FT}{\pi FT} \tag{4}$$

It follows that S(F), the digital counter response to a sinusoidal frequency modulation $A\sin 2\pi Ft$ of the input frequency is:

$$S(F) = A\frac{\sin\pi FT}{\pi FT} \tag{5}$$

In order to estimate the upper bound of error contributed by an interfering magnetic field assume the field to be of the form:

$$\phi_1(t) = \phi_1\sin 2\pi F_1 t \tag{6}$$

From Equation 5, the contribution of the interference to the measured value of the field with a counter is:

$$\phi_1\frac{\sin\pi F_1 T}{\pi F_1 T} < \frac{\phi_1}{\pi F_1 T} \tag{7}$$

For example, to find the effect of the Convair 580 propeller interference (with $\phi_1 = 1\gamma$ and $F_1 = 17$ Hz) one has:

$$\frac{1\cdot\gamma}{\pi F_1 T} \approx \frac{\gamma}{53T} \tag{8}$$

Thus, for a 0.01γ maximum contribution to $\phi_M(t)$, T would have to be approximately 2 seconds; similarly, for a 0.1γ contribution, a T of 0.2 seconds would be sufficient. It follows that, if the magnitude and the frequency of an interference are known, an averaging period T can be selected to reduce its contribution below a specified level.

The uncertainty of count inherent in an electronic digital counter is one count. With a gate or measuring period T and average input frequency $\bar{f}$ the counter will accumulate N counts according to:

$$N = \bar{f} \quad T \tag{9}$$

For a one count uncertainty, the frequency uncertainty is

$$\Delta\bar{f} = \frac{1}{T} \tag{10}$$

If the input frequency is multiplied by a factor K then the frequency uncertainty relative to the input becomes

$$\Delta\bar{f}_{in} = \frac{1}{KT} \tag{11}$$

The corresponding magnetic field resolution is

$$\Delta\phi = \frac{\Delta\bar{f}_{in}}{\alpha} = \frac{1}{\alpha KT} \tag{12}$$

where α is defined by Equation (1).

Thus, an increase in the resolution of frequency measurements may be obtained by multiplying the input frequency by a factor K. For example, to obtain 0.1γ resolution with a 0.1 second gate time T, a K of approximately 29 is required.

The frequency multiplication is achieved, in practice, with a phase locked loop, referred to later as PLL. There is a practical upper limit to the multiplication factor K that can be used. This limit depends on the phase noise and disturbances added by the PLL and on the phase noise of the input frequency (Ref. 8).

Small frequency changes can be measured incrementally with a high resolution frequency-to-voltage converter (Ref. 6) such as a delay line frequency discriminator, later referred to as DLD. In a DLD a periodic input, such as $V(t) = A\sin 2\pi ft$ is delayed by τ by means of a delay line to form $V(t-\tau) = A\sin 2\pi f(t-\tau)$. The delayed input and the non-delayed input are connected to a phase detector. The output of a phase detector is a voltage v_o proportional to the phase difference ψ between the two inputs. The phase difference is given by:

$$\psi = av_o = 2\pi ft - 2\pi f(t-\tau) = 2\pi f\tau \tag{13}$$

The phase is linear with frequency and is a periodic function of angle with a period of 2π; therefore, to each 2π phase change there corresponds a frequency change Δf such that $\Delta f \cdot \tau = 1$ or $\Delta f = \frac{1}{\tau}$. Equation (13) becomes

$$\psi = 2\pi f\tau = 2\pi \frac{f}{\Delta f} = 2\pi(n+\theta) = 2\pi\theta \tag{14}$$

where n is an integer and θ is a fraction of Δf.

A diagram of the DLD is shown in Figure 1A. The input frequency is converted by a pulse shaper into a train of short pulses corresponding to, say, positive zero crossings. These pulses are then delayed in a delay line. The delayed and non-delayed pulses are used to set and reset a flip-flop operating as a

conventional phase detector. Its duty ratio is directly proportional to the phase difference between the two trains of pulses. The output of the flip-flop operates a precision switch, which switches between two reference levels: ground and V_R. The output of the switch is filtered with a low-pass filter. Such a filter will have an effective time constant much larger than the period of the input frequency and it will provide an output v_o proportional to the time average of the input. The output voltage is described by Equation (15) and is shown in Figure 1B:

$$v_o = V_R \frac{t_{on}}{t_{on} + t_{off}} = V_R \theta \tag{15}$$

The resolution of the DLD is very high, limited only by the circuit noise. For example, with a delay of 250 microseconds the Δf is 4000 Hz. By multiplying the input frequency by a factor of 4, a Δf equivalent to 1000 Hz is obtained. A 20 volt peak-to-peak output from the phase detector will then give a scale factor of 10 mV/Hz, permitting the display of 1 Hz as full scale on a strip chart recorder.

The linearity and repeatability of a prototype system designed at the Flight Research Laboratory was found to be better than 0.1%. The design details of such a delay line frequency discriminator (DLD) are given in Appendix A.

3.0 TOTAL MAGNETIC FIELD GRADIENT MEASURING TECHNIQUES

The magnitude of earth's total magnetic field is a scalar field; its gradient is a vector. To determine the gradient uniquely it is necessary to know its position, its magnitude and its direction. The co-ordinates of the gradient vector's origin are longitude, latitude and altitude obtained from navigational or position finding data. The gradient can be considered to be a vectorial sum of its components. Each component, which is a spatial partial derivative, can be approximated by a simultaneous differential measurement of the magnetic field at two proximate positions located on a line in a desired direction. The minimum separation of the measuring sensors is dictated by the accuracy desired for the gradient and by the resolution with which a magnetic field difference can be measured. For example, on an aircraft the sensors can be mounted on wing tips or on the top and on the bottom of the vertical tail fin. Often, it is only required to measure one component e.g. the vertical, commonly referred to as the "vertical gradient". The following discussion will treat the measurement of only one component of the gradient, with the understanding that the same measuring techniques can be applied to other components.

As explained earlier, the measurement of the magnetic field difference with optically pumped magnetometers is reduced to the measurement of frequency differences.

It should be mentioned that, usually, only small frequency differences are measured because the magnetometers are subjected to similar but slightly unequal inputs.

The principle of a typical digital frequency difference measuring circuit is shown in Figure 2. In order to reduce operating frequencies for the PLL's and for the counters, a down-conversion can be used as shown in Figure 3. Both channels for f_1 and f_2 are subject to identical processing.

Referring to Figure 3, the output $\bar{D}_n$ from the subtracting circuit will be:

$$\bar{D}_n = \int_{nT}^{(n+1)T} D[f_1(t) - f_o]dt - \int_{nT}^{(n+1)T} K[f_2(t) - f_o]dt = K\int_{nT}^{(n+1)T} [f_1(t) - f_2(t)]dt$$

$$= KT(\bar{f}_1 - \bar{f}_2) \tag{16}$$

The gating period T and multiplication factor K will be selected to satisfy a given interference rejection and a resolution condition. At the same time T has also to be chosen to achieve a desirable sampling rate.

Reference 9 describes an airborne earth's total field measuring system which evolved into an airborne gradiometer described in Reference 10. The gradiometer design is based on the frequency difference measuring system, as shown in Figure 3.

The accuracy to which a magnetic field can be measured will depend on interference level, system noise, timing accuracy and the ambiguity of one count in the digital counter.

An analog frequency difference measuring system is shown in Figure 4. It consists of two DLD's, the outputs of which are subtracted by means of two operational amplifiers. A scale factor correction is required to compensate for different delays in each delay line. The accuracy of such a circuit depends, to a large extent, on the quality of the DLD's used.

Another frequency difference measuring system is shown in Figure 5. This arrangement consists of a mixer circuit with a low-pass filter to provide the frequency difference, followed by a digital measuring system in the form of a universal counter-timer (e.g. Philips PM6650 or similar) capable of measuring frequency and period or multiple periods. According to the identity

$$\cos(\omega_1 t + \psi_1)\cos\omega_2 t = \tfrac{1}{2}\left\{\cos[(\omega_1 + \omega_2)t + \psi_1] + \cos[(\omega_1 - \omega_2)t + \psi_1]\right\} \qquad (17)$$

the output of the mixer will contain the sum and the difference of ω_1 and ω_2. After a low-pass filter the mixer output V_o will be:

$$V_o \cong \tfrac{1}{2}\cos[(\omega_1 - \omega_2)t + \psi_1] = \tfrac{1}{2}\cos(2\pi\Delta f t + \psi_1) \qquad (18)$$

It can be observed that the mixing process preserves the phase ψ_1, which may have a large random component $\hat{\psi}(t)$ (e.g. with a Gaussian distribution). The frequency difference, or beat frequency, is usually small, and to measure it directly may require too much time. For example, to measure directly 1 Hz to a 0.1 Hz accuracy, it would take at least 10 seconds, severely restricting the sampling rates available. Furthermore, with this arrangement multiplication by a factor K to increase the resolution in a given measuring period T is difficult when the frequency ratio $\frac{\Delta f\ max}{\Delta f\ min}$ is large. Also a large ratio complicates the design of the PLL. Another difficulty with this arrangement arises from the ambiguity which exists as to whether $f_1 > f_2$ or $f_2 > f_1$; thus, the Δf is "unsigned".

Period measurements present similar difficulties. It takes a long time to measure periods of low frequencies because the period tends to infinity as the frequency approaches zero. The phase noise $\hat{\psi}(t)$ makes the period readings variable. To reduce this variability an average of N periods must be taken, further increasing the time of measurement and adversely affecting the sampling rate.

To overcome these difficulties, it is desirable to offset Δf by a reference frequency f_R, achieved, for example by adding f_R to f_1 and forming $(f_1 + f_R) - f_2 = f_R + \Delta f$.

The frequency offset or up-conversion of Δf by f_R is ultimately achieved by the multiplication (mixing) of two frequencies, a process which does preserve the phase angle ψ of Δf (see Eq. (17)), provided f_R has a negligible phase noise. Assuming $\Delta f \ll f_R$, during a measuring period T, there are N phase samples, where $N \cong Tf_R$. Selecting f_R and T, as large as practical, the number of phase samples N can be optimized to obtain phase averages of a large number of samples in a reasonable measuring period T. A specialized circuit, referred to as a frequency difference translator (FDT) has been developed by the author at the Flight Research Laboratory for this purpose. The f_R was selected to be 111.111 kHz although it could have been selected anywhere between 100 kHz to 200 kHz. Choosing such a high frequency offset makes the output compatible with existing instrumentation designed to operate in the 100 kHz to 200 kHz range, (corresponding to the cesium-vapour magnetometer output in the earth's magnetic field).

4.0 FREQUENCY DIFFERENCE TRANSLATOR

The block diagram of the frequency difference translator is shown in Figure 6 and the design details are given in Appendix B.

The principle of operation is that of subtracting the output frequencies of two phase lock translation loops. The output frequency f_o can be expressed as follows:

$$f_o = \left(f_1 + \frac{f_R}{p}\right) - \left(f_2 + \frac{f_R}{q}\right) = f_R\left(\frac{q-p}{pq}\right) + f_1 - f_2 = kf_R + \Delta f \qquad (19)$$

The phase lock translation loops used can be easily modified to generate $mf_1 + f_R$ as shown in Figure 7, resulting in the frequency difference translator having an output:

$$f_o = Bf_R + m\Delta f \qquad (20)$$

A measuring system is made up of a frequency difference translator supplied with a stable reference frequency f_R followed by a delay line frequency discriminator and a low-pass filter, the output of which can be displayed on a strip chart recorder or digitized for further processing. Such a measuring system is shown in Figure 8.

To measure a frequency difference $f_2 - f_1$, exceeding the linear frequency span or period of the DLD's output, a frequency counter may be added (the DLD's output period is shown in the Figure 1B and denoted Δf, not to be confused with $\Delta f = f_2 - f_1$ used presently). Such a counter will have to measure the approximate output frequency, f_o to determine by how many DLD frequency periods it differs from kf_R (Eq. (19)).

The accuracy of the system is limited only by the noise generated in associated mixers and oscillators. In the prototype design a resolution of the order of 0.01 Hz in a 5 Hz bandwidth has been achieved and a further improvement in resolution to 0.001 Hz in reduced bandwidth can be anticipated.

The advantages of such a system are mainly stability, high sensitivity, simplicity and speed of measurement not previously attainable. It is essentially a system operating in real-time in contrast to the system using digital counters which provides the data with a time delay.

One of the important differences between the FDT followed by a DLD, and the system operating on the principle of digital counting and subtracting, is that the former system is linear in frequency and continuous while the latter is a sampled data system. The output of the DLD is a voltage proportional to the input frequency. This voltage can be processed with a filter to suit particular requirements. For example, a high order low-pass filter will provide better rejection of known high frequency interferences than a digital counter (see Eq. (5)). Such a filter will reduce the unwanted effects of frequency modulation of the input frequency present in the DLD's output.

5.0 CURRENT AND PROPOSED APPLICATIONS OF THE FREQUENCY DIFFERENCE TRANSLATOR

The most obvious application is that of measuring the earth's total magnetic field gradient and of variables associated with the gradient. This instrumentation was developed primarily for this application.

There is another class of FDT applications, not immediately obvious, resulting from the elimination (subtraction) of magnetic disturbances originating from far away sources such as magnetic micropulsations, which are time varying, quasi-random changes in the earth's magnetic field and which can be of several gammas in magnitude. This is of paramount importance in obtaining a magnetically

quiet environment for experimental purposes. In such applications, one magnetometer is subjected to an external stimulus while the second one, located as close as possible to the first one, acts as a "base" or reference. Both magnetometers are subject to the same external "noise" which is eliminated by subtraction, allowing for better detectability of small magnetic changes. With a single magnetometer, a response to a small change in the magnetic field can be buried in the "noise". It should be mentioned that sensitive magnetic measurements of this kind should be made in a nonmagnetic environment removed from man-made magnetic interferences (which may be from a close-by source, thus generating large gradients).

Magnetometer stability can be measured by placing two similar magnetometers a few feet apart and recording the time history of the frequency differences. An assumption is made that their variations are random and uncorrelated and that both contribute equally to the measurement.

Temperature sensitivity can be measured by subjecting one of the two magnetometers to temperature variations, while the other is kept at constant temperature.

Turn-on retrace can be studied by turning one magnetometer on and off periodically and observing whether or not the output returns to the same value.

Also, variations of the output as a function of the magnetometer position can be measured with respect to the earth's magnetic field vector. Figure 9 shows measured changes of a magnetometer output as a function of rotation about its own axis. Figure 10 shows variations in the output as the magnetometer is tumbled in a vertical plane parallel to the North-South bearing. Two "working" sectors are clearly visible, that is, the angular sectors to which correspond small output changes. Figure 11 shows the variation when the magnetometer is rotated in a horizontal plane. Note that the data were obtained from a compensated, "very good" magnetometer. Other magnetometers may exhibit much larger variations. The above mentioned measurements will allow assessment of the possible system accuracy using a given magnetometer. Furthermore, some corrections can be introduced once a model is known. This can be particularly useful in a strapped-down configuration; that is, with the magnetometer in a fixed position with respect to the aircraft, rather than with the self orienting type, (Ref. 11).

The time required to make the measurements in Figures 9, 10 and 11 is relatively short, of the order of a few minutes; furthermore, the measurements repeated several times showed a very good consistency of the results. Previous attempts at similar measurements, using a counter-timer connected as shown in Figure 5, resulted in lengthy procedures requiring considerable averaging time for each data point and the repeated measurements showed considerable scatter.

The frequency difference translator could be used for the magnetic clean-up of the magnetometer carrying vehicle, in our case, an aircraft. Assorted electrical systems and magnetic moving inclusions generate a man-made magnetic noise which can reduce the accuracy of measurements; therefore, it is essential to reduce this noise to a minimum.

The clean-up procedure consists of placing one magnetometer away from the aircraft while the other is placed at various locations in or on the aircraft. The aircraft power is then turned on and various electrical systems are energized. The interference from these systems can now be observed and minimized by twisting power leads, using a "one only" grounding point, avoiding current loops, using magnetic shielding etc. Similarly, magnetic inclusions in moving parts, e.g. magnetic steel bolts in ailerons, can be detected and when possible replaced by non-magnetic equivalents.

In general, the knowledge of an objectionable interference source allows for a suitable counteraction. The aim is to have in the aircraft only fixed and constant magnetic moments whose behaviour can be modelled mathematically. Use of the frequency difference translator allows measurement of influences, which may later be eliminated, leading ultimately to a magnetically quiet aircraft.

6.0 CONCLUSIONS

The techniques used to measure total magnetic field gradient using optically pumped cesium vapour magnetometers have been discussed. The discussion has focussed on the processing of the output frequency differences. There are other transducers and magnetometers with an output of frequency proportional to the input variable, to which the same processing techniques will apply.

There are two basic methods of measuring frequency: one relying on counting, the other on frequency-to-voltage conversion. With the counting method the output is time-delayed, since it always takes time to count and is only available in sampled form, while with the frequency-to-voltage conversion method the output is continuous and allows for real-time processing.

To facilitate the measurement of small frequency differences the Flight Research Laboratory has developed a frequency difference translator which, with the delay line frequency discriminator, makes a very versatile instrument providing precise readings. The output of such a system is a voltage which is linear with respect to frequency or frequency difference. This voltage can be processed with a frequency filter. The filter can be active, of high order, and can be designed to accommodate specific requirements such as rejection of known interferences.

The wide selection of readily available "standard" designs for active frequency filters and the precision in forming a frequency difference are the two main advantages of the frequency difference translator system over all the other systems discussed.

The frequency difference translator should find many applications e.g. in frequency comparison, in frequency deviation monitoring and in providing continuous differential readings of other transducers with a variable frequency for an output. At the time of this writing, the frequency difference translator was a new development and many of these applications have yet to be explored.

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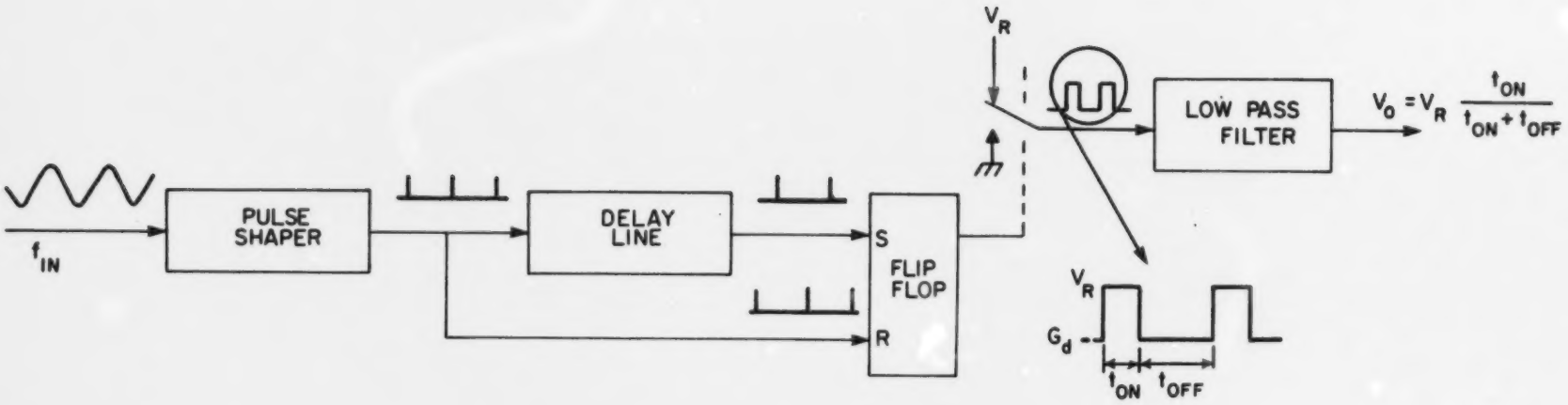

FIG. 1A: DELAY LINE FREQUENCY DISCRIMINATOR

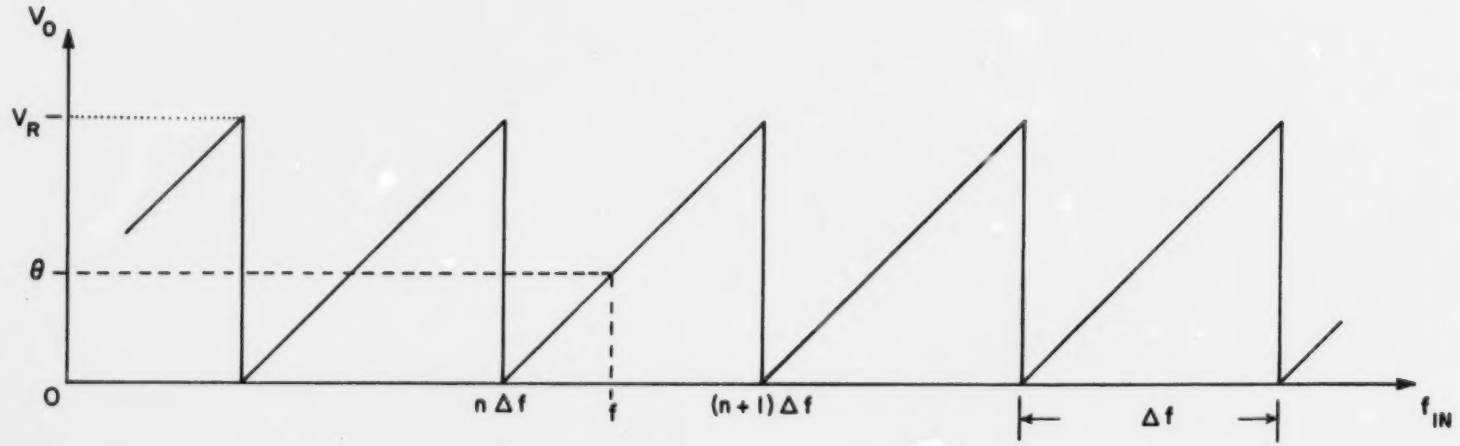

FIG. 1B: OUTPUT OF DELAY LINE DISCRIMINATOR

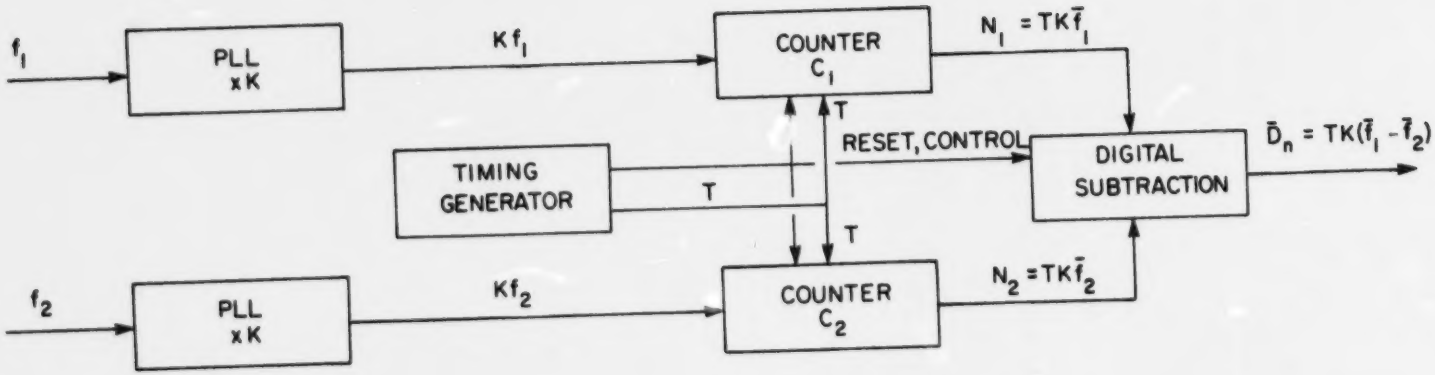

FIG. 2: DIGITAL FREQUENCY DIFFERENCE MEASURING CIRCUIT

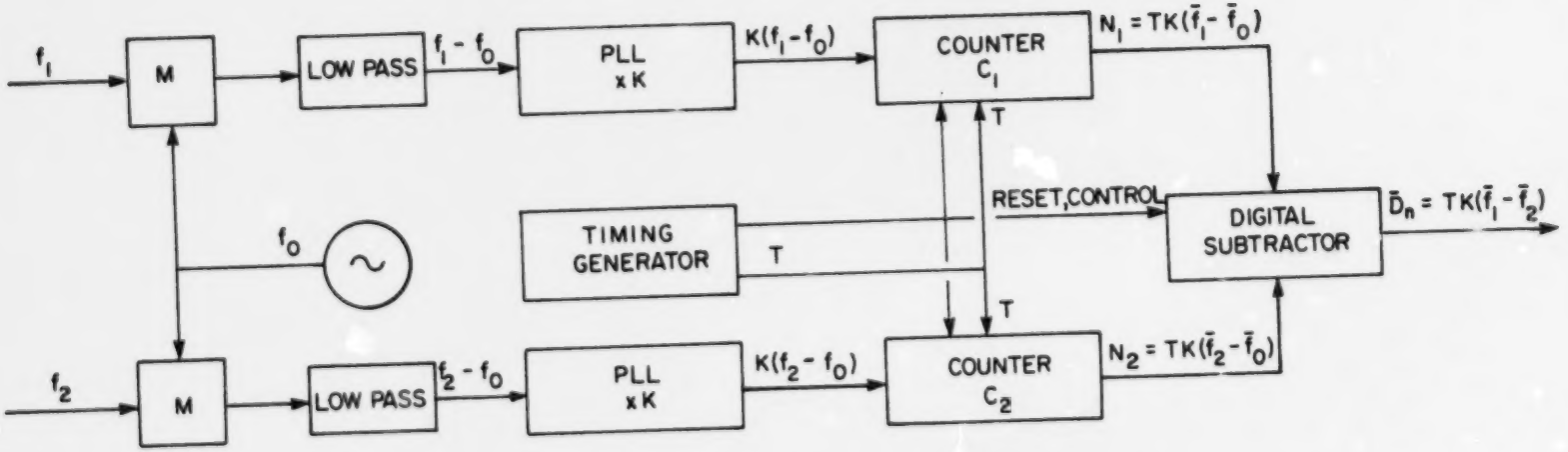

FIG. 3: DIGITAL FREQUENCY DIFFERENCE MEASURING CIRCUIT WITH DOWN CONVERSION

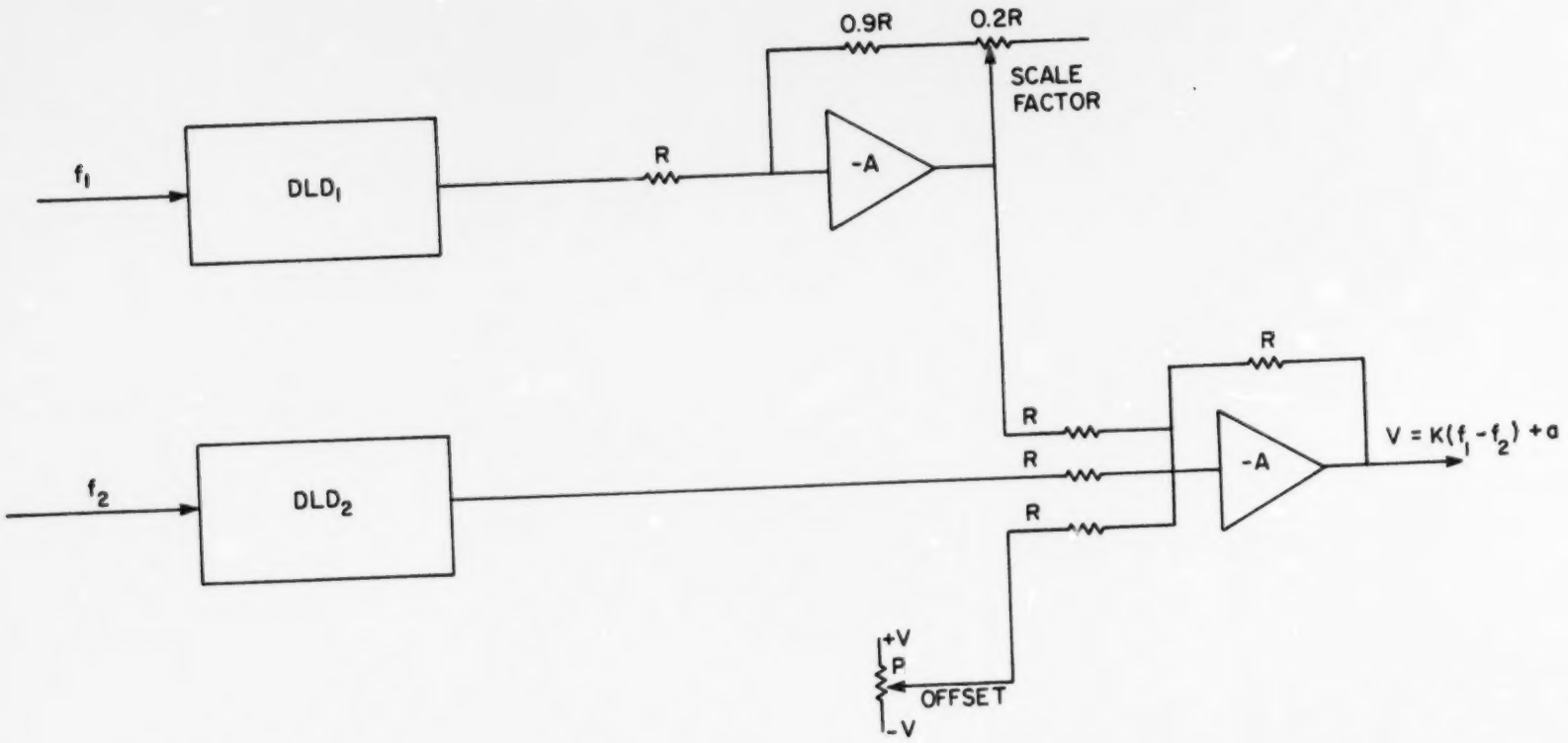

FIG. 4: ANALOG FREQUENCY DIFFERENCE MEASURING CIRCUIT

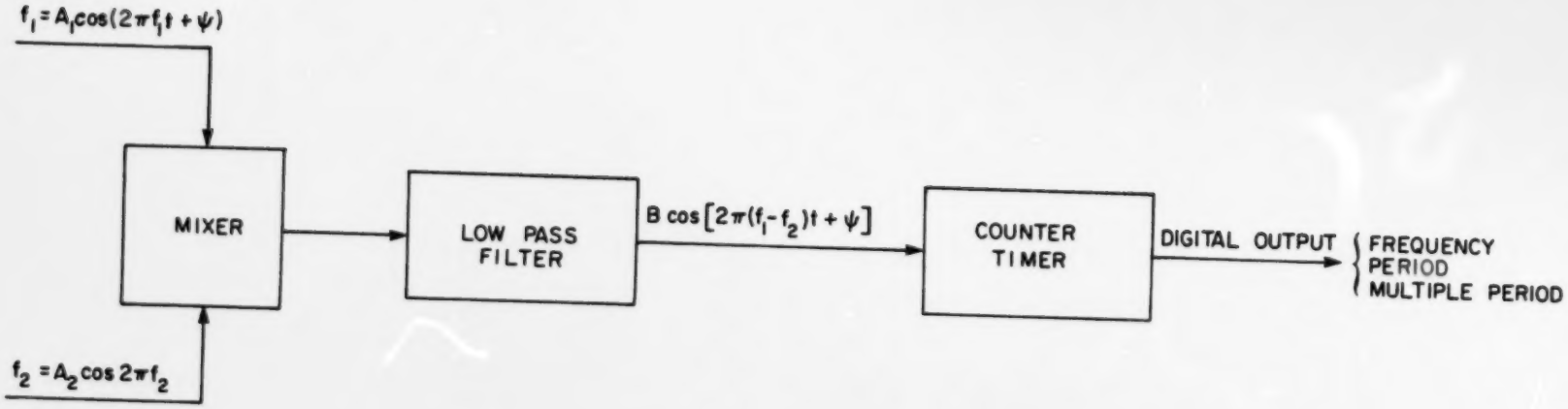

FIG. 5: FREQUENCY DIFFERENCE MEASURING CIRCUIT USING A MIXER

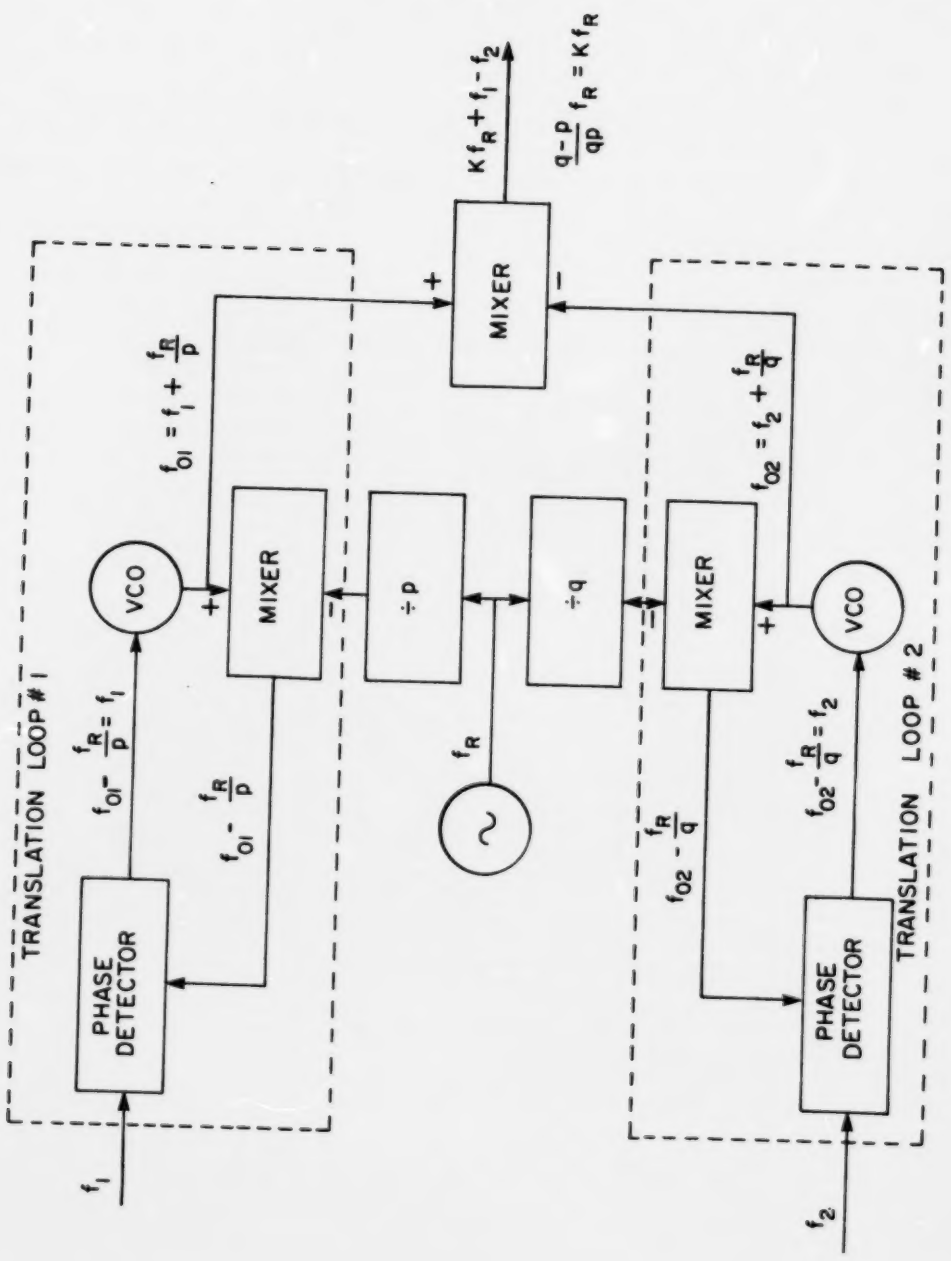

FIG. 6: FREQUENCY DIFFERENCE TRANSLATOR

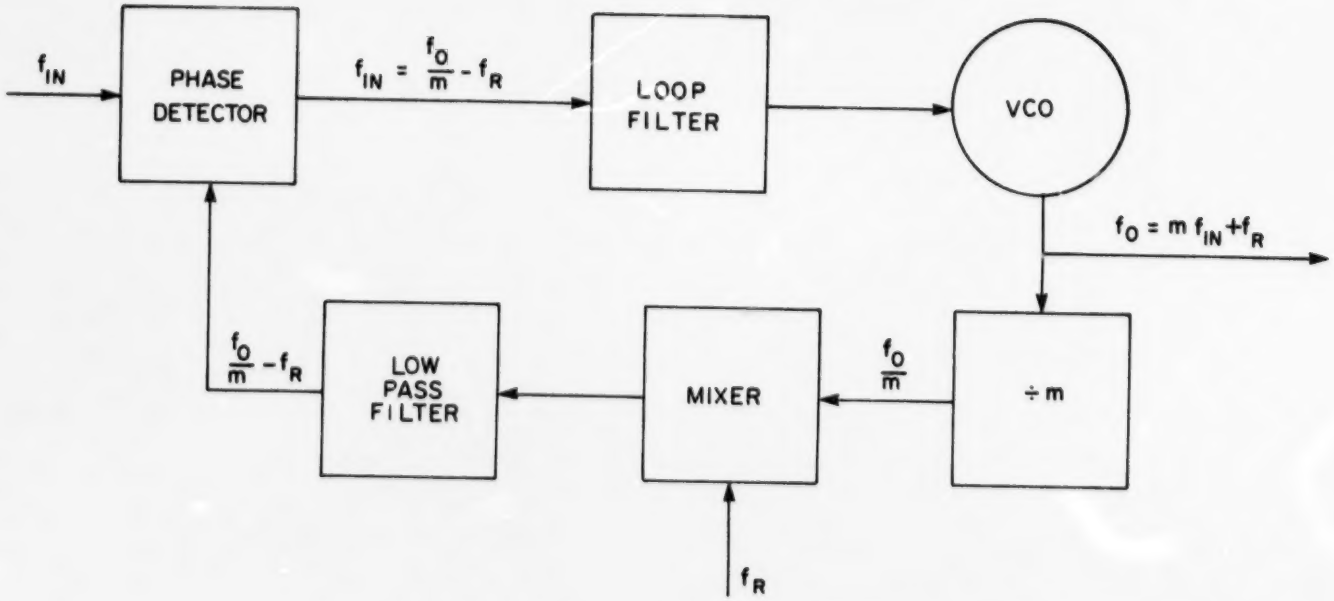

FIG. 7: PHASE LOCK TRANSLATION LOOP WITH A DIVIDER

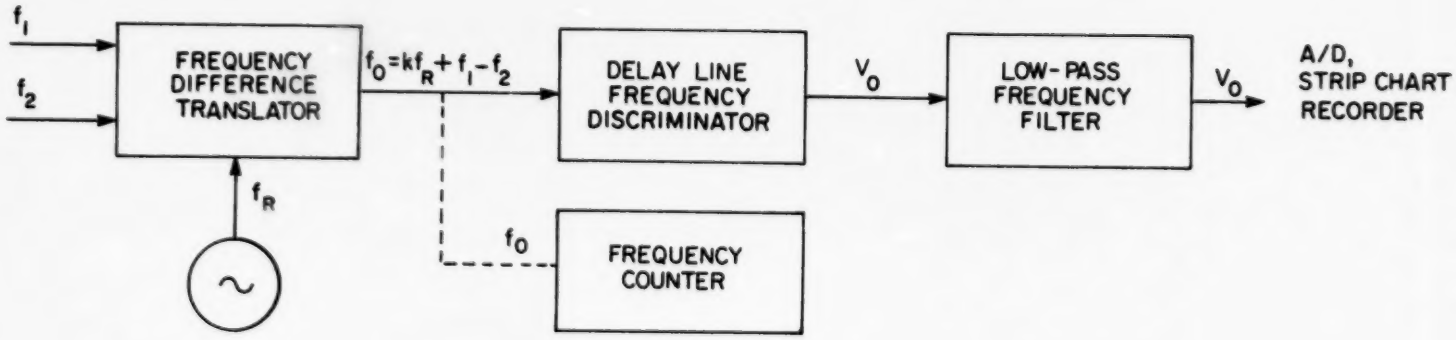

FIG. 8: FREQUENCY DIFFERENCE MEASURING SYSTEM USING FREQUENCY DIFFERENCE TRANSLATOR

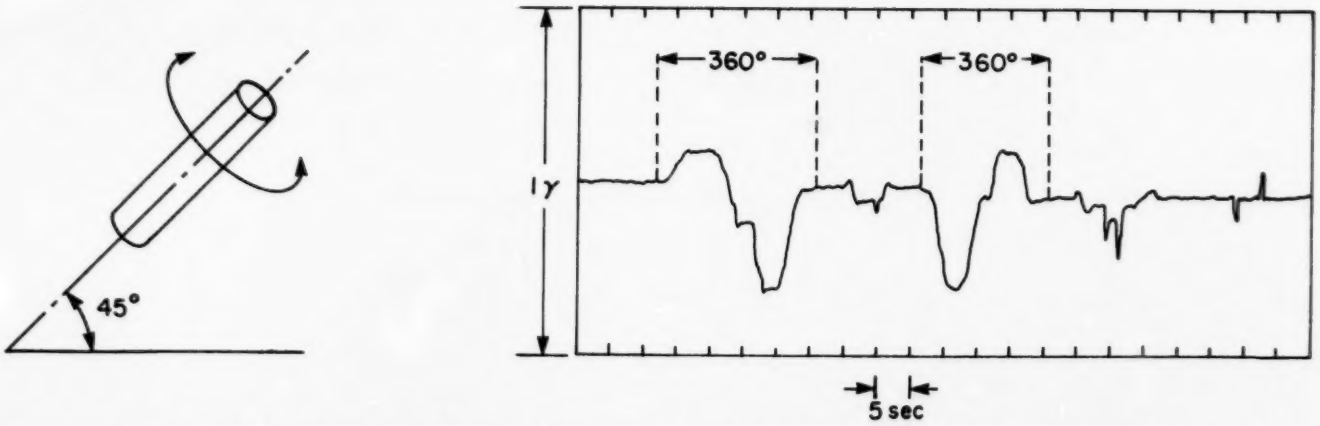

FIG. 9: MAGNETOMETER OUTPUT AS A FUNCTION OF ROTATION ABOUT ITS OWN AXIS

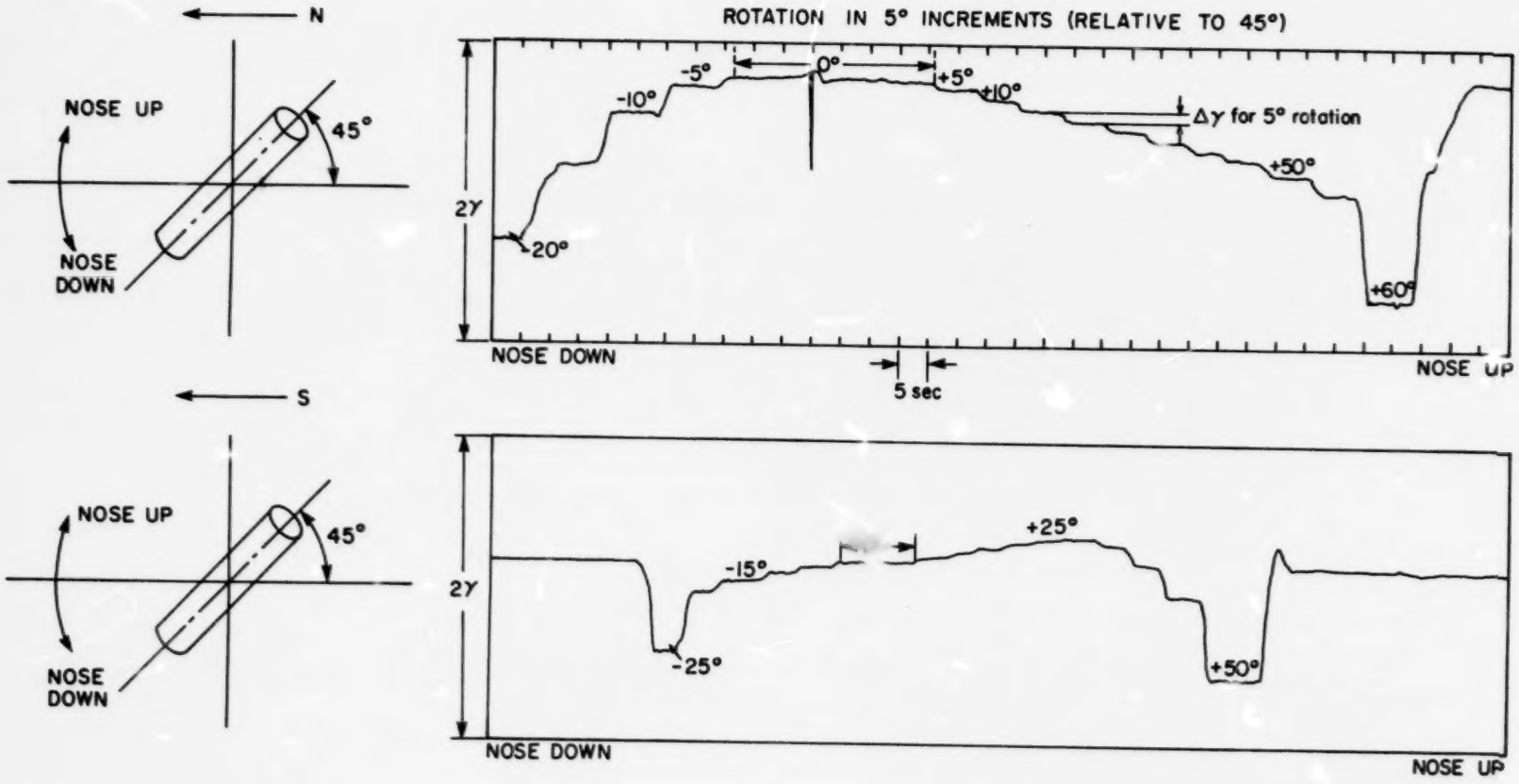

FIG. 10: MAGNETOMETER OUTPUT AS A FUNCTION OF ROTATION IN VERTICAL PLANE

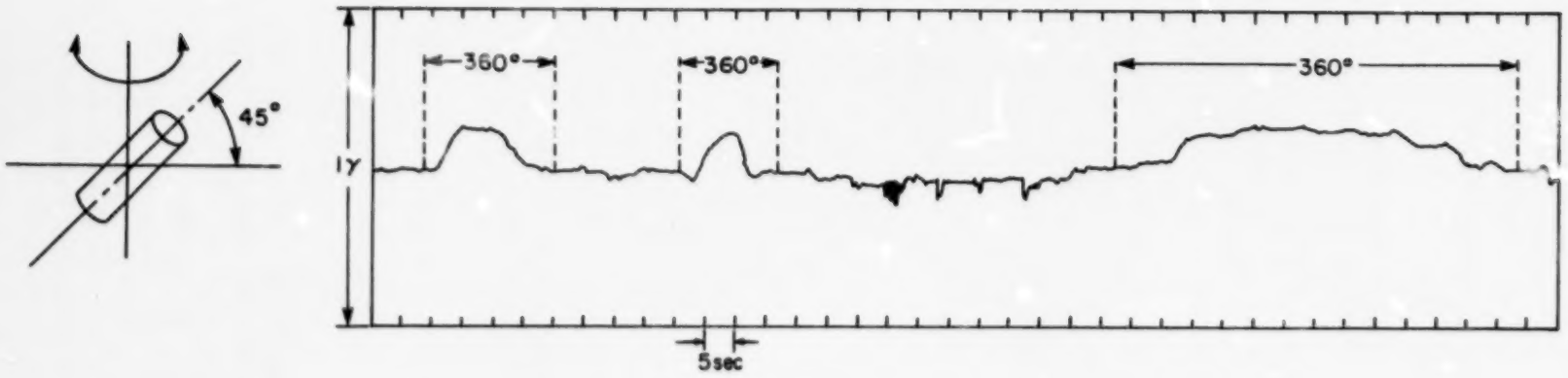

FIG. 11: MAGNETOMETER OUTPUT AS A FUNCTION OF ROTATION IN HORIZONTAL PLANE

APPENDIX A

DELAY LINE FREQUENCY DISCRIMINATOR DESIGN CONSIDERATIONS

The principle of operation of the DLD has been described previously in the text; this note serves to give some engineering information which would be useful in designing a similar device.

A simplified schematic of a DLD developed by the Flight Research Laboratory is shown in Figure A1. The phase lock loop (PLL) at the input is constructed from a phase frequency detector (MC4344), a dual voltage-controlled multivibrator (MC4324) and a divided-by-four circuit (SN54LS73). Technical data for similar PLL designs is contained in Reference A-1.

The "one-shot" (O.S.) or monostable multivibrator (SN54121) is followed by a buffer B to drive the 250 μs delay line. The buffer consists of four sections of SN54128 wired in parallel. The pulse width from the "one-shot" (100 nS) and buffer will depend on the bandwidth of the delay line used. The delay line used was an ultrasonic type made of glass, with a very low temperature coefficient of delay. The output of the delay line is processed with a high speed comparator C (LM161).

The combination of inverters and NAND gates forms an edge-triggered flip-flop part of a conventional phase detector. The flip-flop is set by the delayed pulse and reset by the non-delayed pulse. The switch SW and the Exclusive-OR gate in the reset path provide for flip-flop reset with either the leading edge or the trailing edge of the input square-wave, altering the flip-flop's duty ratio by one half of a period. The outputs of the flip-flop after level translation circuits L (SN 5407), operate precision switches S (CD 4050), switching between ground and a reference voltage level V_R. The operational amplifier A_1 with its associated low-pass filters subtract the average values of the complementary wavetrains generated by the flip-flop and switches S. The output of A_1 is a voltage proportional to the input frequency (phase), repeating every Δf, which in this design is 1000 Hz. Note that the switch SW changes the output voltage by an amount equivalent to 500 Hz.

The operational amplifier A_2 is also used as a low-pass filter. While A_1 has a cut-off frequency of the order of a few kHz, the response of A_2 is usually tailored to specific requirements, e.g. 60 Hz interference rejection.

The potentiometer P is useful to off-set the output; for example, it may be required to set the output voltage close to zero in order to display the output at a high sensitivity on a strip-chart recorder.

The glass delay lines are available commercially with delays from a few microseconds to 4 milliseconds. These delay lines, when used in a delay line frequency discriminator, will provide a wide range of Δf's, which can be further extended by either multiplying or dividing the input frequency.

The limiting factor is the speed with which the switch S will operate. In the prototype design, the maximum frequency of operation is of the order of 1 MHz.

REFERENCE

A-1 Motorola Staff — *Phase Locked Loop Systems.* Motorola Semiconductors Products Inc., Second Edition, Aug. 1973.

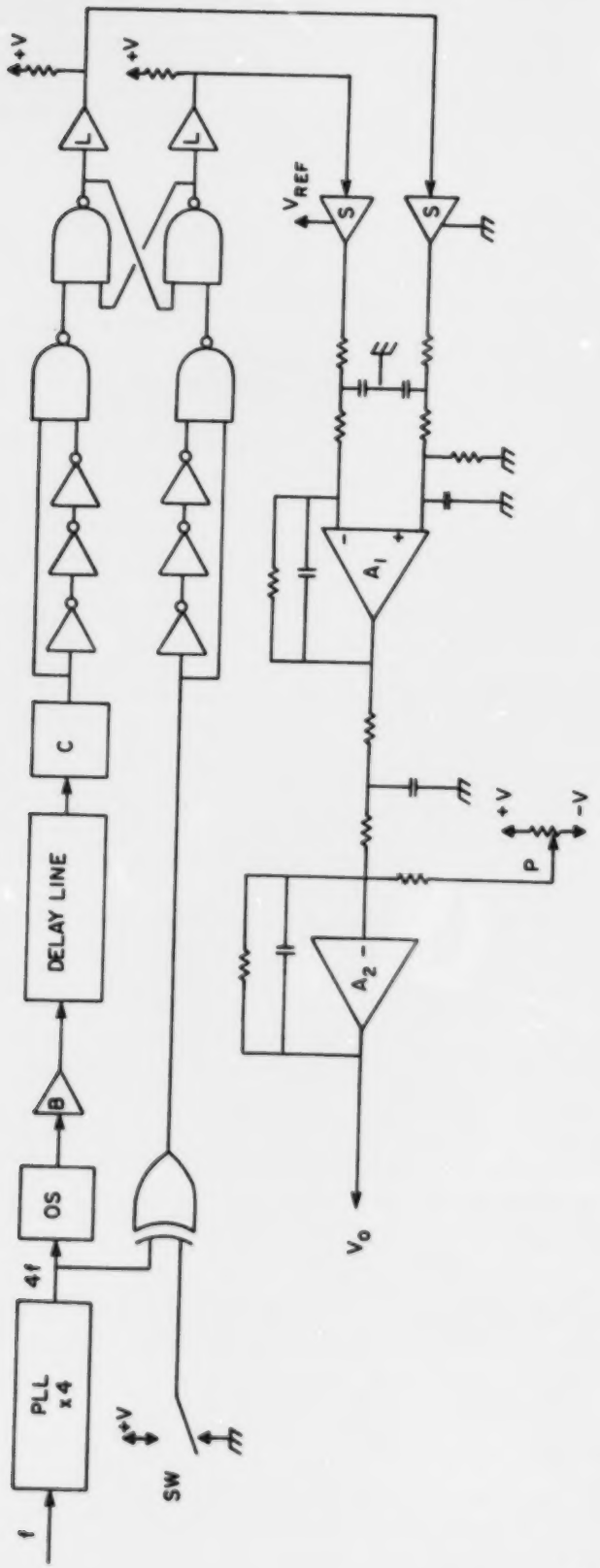

FIG. A1: SIMPLIFIED DIAGRAM OF DELAY LINE FREQUENCY DISCRIMINATOR

APPENDIX B

FREQUENCY DIFFERENCE TRANSLATOR

A frequency difference translator is built from two basic translation loops as shown in Figure 6 of the main text. A simplified schematic of a translation loop is shown in Figure B1.

There are two critical components in the translation loop on which the success of operation depends. These are the low-noise voltage controlled oscillator or VCO, and a mixer. Careful selection of mixer frequencies is also important. A reasonably high ratio of f_o to $f_o - f_2$ has been selected to reduce higher frequency products coinciding with the desired output $f_o - f_2$ (Ref. B-1). With the selected frequencies the interfering products are of 9th order or higher. The selected frequencies are 100 kHz $\leq f_1 \leq$ 200 kHz, f_2 = 1 MHz and 1.1 MHz $\leq f_o \leq$ 1.2 MHz for loop 1; and f_2 = 1.111 MHz and 1.211 MHz $\leq f_o \leq$ 1.311 MHz for loop 2. One of the essential features of the oscillator is its low phase noise. C_1, C_2, D_1, D_2 and R_1 form a voltage doubler in an auxiliary feedback loop which biases the FET transistor into a linear operating range. The low-pass filter which follows the mixer (MC 1596) is a fifth order Butterworth with an approximate cut-off freqency of 200 kHz.

The output of the translation loop is $f_o = f_1 + f_2$. The f_2's for the two translation loops have been selected to be $\frac{10\text{ MHz}}{10}$ and $\frac{10\text{ MHz}}{9}$ respectively. For the 10 MHz source any temperature compensated crystal oscillator (TCXO) can be used. Its desired short time frequency stability is approximately 1×10^{-7}. The division is achieved by conventional digital transistor-transistor logic (TTL). The subtraction circuit, between the two translation loop outputs, as shown in the Figure 6 of the main text is achieved with a mixer (MC 1596 or LM 1596) followed by a fifth order low-pass Butterworth filter, similar to that used in the translation loop.

REFERENCE

B-1 Kroupa F.V. *Frequency Synthesis.* John Wiley & Sons, 1973.

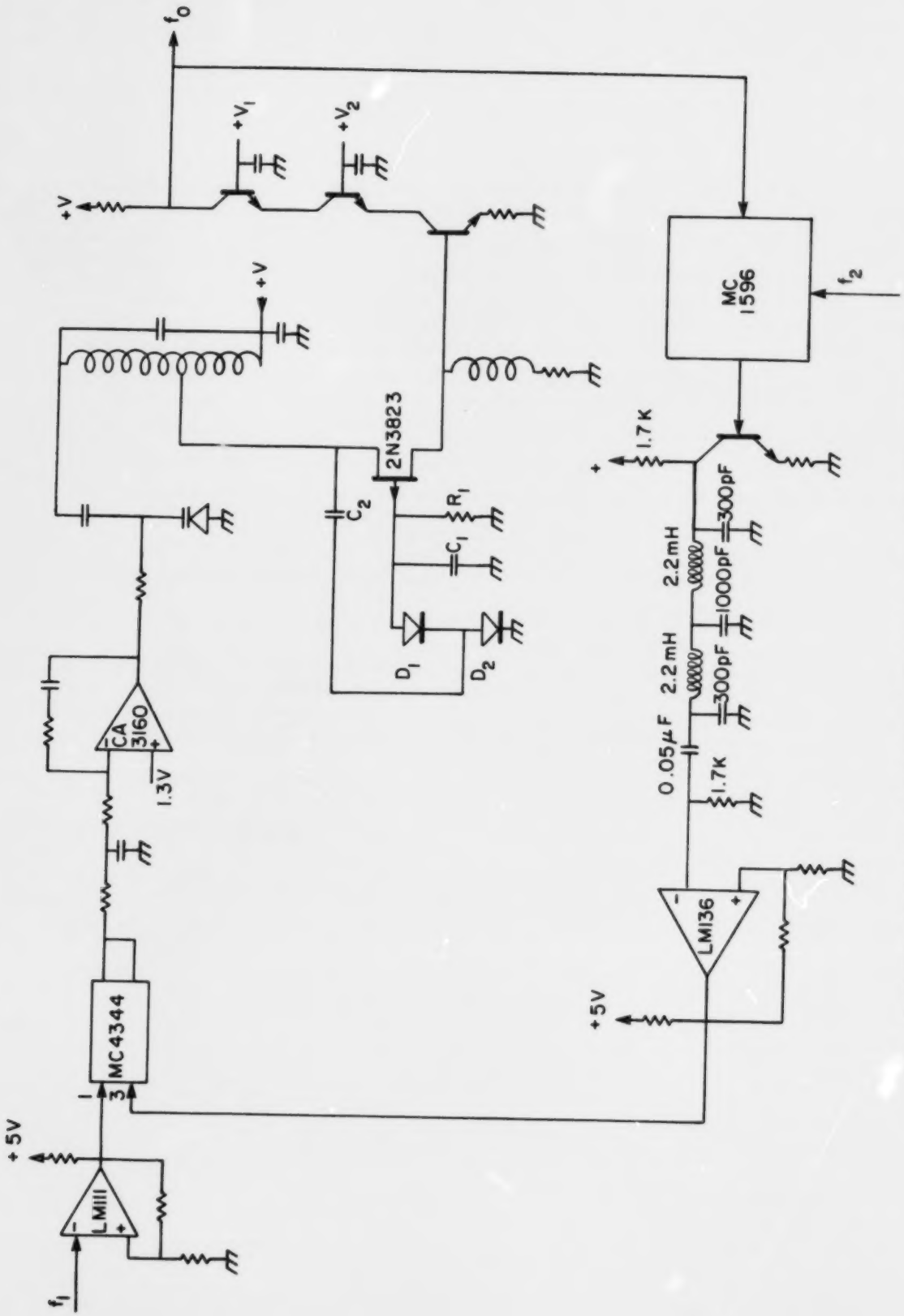

FIG. B1: SIMPLIFIED DIAGRAM OF FREQUENCY TRANSLATION LOOP

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THE FREQUENCY DIFFERENCE TRANSLATOR: AN IMPROVEMENT IN AIRBORNE TECHNIQUES FOR MEASURING THE EARTH'S TOTAL MAGNETIC FIELD GRADIENT.
Barszczewski, A. September 1982. 30 pp. (incl. figures and appendices).

Improvements in Airborne techniques for the measurement of the earth's total magnetic field gradient using optically pumped magnetometers are discussed. These techniques involve fast and accurate measurements of small frequency differences and are applicable to a large class of transducers which have for an output a frequency varying with the input variable. Description and design details are given of a newly developed analog instrument which allows a continuous read-out of frequency difference with a resolution of better than 0.01 Hz.

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2. Geomagnetism
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